文献1

Low-IF Topologies for High-Performance Analog Front Ends of Fully Integrated Receivers

Jan Crols and Michiel S. J. Steyaert, *Senior Member, IEEE*

***Abstract*— When it comes to integratability, the zero-intermediate frequency (IF) receiver is an alternative for the heterodyne or IF receiver. In recent years, the zero-IF receiver has been introduced in several applications, but its performance cannot be compared to that of the IF receiver yet. This lower performance is closely related to its baseband operation, resulting in filter saturation and distortion, both caused by DC-offsets and self-mixing at the inputs of the mixers. The low-IF receiver has a topology which is closely related to the zero-IF receiver, but it does not operate in the baseband, only *near* the baseband. The consequences are that, as for the zero-IF receiver, the implementation of a low-IF receiver can be done with a high degree of integration, however, its performance can be better. In this paper, the fundamental principles of the low-IF receiver topology are introduced. Different low-IF receiver topologies are synthesized and fully analyzed in this paper. This is done by applying the complex signal technique—a technique used in digital applications to the study of analog receiver front ends.**

I. Introduction

THIS PAPER deals with the design of receivers for telecommunication systems in which phase or frequency modulation is used. It is investigated how these receivers can be realized in a highly integrated way and still show a good quality of signal reception. This is not trivial. Today, most receivers work with an intermediate frequency (IF), and a lot of their components still need to be discrete. Alternative topologies, mainly the zero-IF topologies, can be realized highly integrated, but they tend to have a significantly poorer performance than the discrete realizations.

A typical receiver consist of two important parts which perform its main operations: the downconversion and the demodulation part. In the downconversion part, the wanted signal is filtered and separated from its neighbors, and it is converted from its carrier frequency to a frequency more suited for the demodulator. Due to the high-performance specifications which are required from the downconverter, its implementation is always fully analog. The demodulation is done on a lower frequency. In more and more applications, this demodulation is done in a digital signal processor (DSP). A DSP allows for the use of complicated modulation schemes and complex demodulation algorithm. The result is an overall quality improvement and a high degree of integration. In this paper, only the design of the analog downconversion part

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The authors are with the Katholieke Universiteit Leuven, ESAT-MICAS, B-3001 Heverlee, Belgium.

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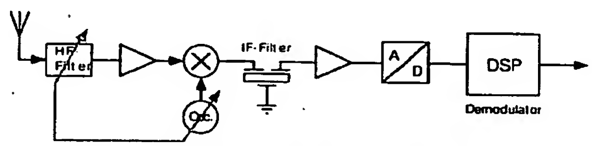

Fig. 1. Schematic representation of the IF receiver topology.

of the receiver is discussed. It will be presumed that the demodulation is preferably done with a DSP at baseband. The design of a downconverter will be discussed accordingly.

There are two main types of receivers used in telecommunication systems based on phase or frequency modulation: the heterodyne and the homodyne receiver. The difference between them is in whether or not an IF is used. This has great implications on their topology, possibilities, and performance. Heterodyne and homodyne receivers are often called IF and zero-IF receivers, respectively. In this paper, the latter term will be used.

The main advantage of a zero-IF receiver over an IF receiver is the very high integration level that can be achieved. Although full integration is very important for cost reduction, the use of zero-IF receivers was very limited in the past due to the poor performance compared to IF receivers. It is only nowadays that one begins to use zero-IF receivers in systems based on digital communications. In these systems, a lower performance can be accepted in exchange for the higher degree of integration and the ease with which a zero-IF receiver can be combined with a DSP for the baseband demodulation of the digital signal. The recently introduced new wireless digital telecommunication services, like GSM and DECT, are examples of such systems [1], [2].

In this paper, all the advantages and disadvantages concerning integratability and performance of IF and zero-IF receivers are discussed. Starting from the discussion of these two topologies, new receiver topologies, based on the use of a low IF, situated at a few hundred kilohertz, are introduced. It is shown how the advantages of IF and zero-IF, a high performance and a high degree of integration, can be combined. For this purpose, the complex signal technique is introduced to this field of analog circuit design. The complex signal principle brings a higher level of abstraction to, and therefore a clearer insight into, the analysis of multipath signal processing systems. The performance of multipath systems relies, however, on a very good matching, which can only be perfect in digital systems. This paper demonstrates how the complex signal analysis technique can be used to get a better

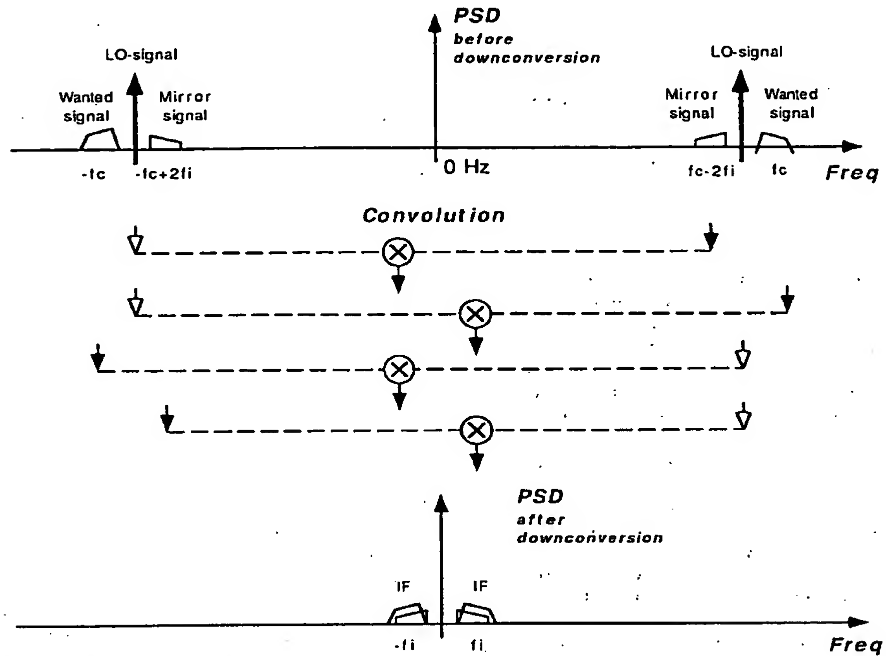

Fig. 2. The downconversion scheme in an IF receiver.

insight into the effect of mismatch in analog multipath signal processing circuits such as zero- and low-IF receivers.

II. Integrated Receivers

A. IF Receivers

IF receivers have been in use for a long time, and their way of operating is very well known [3]. In Fig. 1, the schematic representation of the IF receiver topology is given. The wanted signal is downconverted from its carrier to the IF by multiplying it with a single sinusoidal signal. It can be demodulated on this intermediate frequency or it can be further downconverted after filtering. The main disadvantage here is that apart from the wanted signal, an unwanted signal (called the mirror frequency) is downconverted to the IF. This is illustrated in Fig. 2. When the wanted signal is situated on f_c, the mirror frequency is at $f_c - 2f_i$. This is illustrated with (1). The mixer gives a frequency component on f_i for $f_x = f_c$ and $f_x = f_c - 2f_i$. The antenna signal $a(t)$ is in this case a phase modulated signal and equal to $\cos[2\pi f_x t + m(t) + \phi]$.

$$
\begin{aligned}
&a(t) \times \sin[2\pi(f_c - f_i)t + \varphi] \\
&\quad = \cos[2\pi f_x + m(t) + \phi] \times \sin[2\pi(f_c - f_i)t + \varphi] \\
&\quad = \tfrac{1}{2}\{\sin[2\pi(f_x + f_c - f_i)t + m(t) + \phi + \varphi] \\
&\qquad - \sin[2\pi(f_x - f_c + f_i)t + m(t) + \phi - \varphi]\}. \qquad (1)
\end{aligned}
$$

The mirror frequency has to be suppressed before it is mixed down to the IF. This is done by means of a high-frequency (HF) filter. Such an HF filter can only be realized if f_i is high enough because the wanted signal (on f_c) must be relatively far away from the mirror frequency. Even when the ratio f_c/f_i is as small as 10, the specifications for the HF filter are very severe. The HF filter must have a very high Q (up to 50 and more), and it must have an order which is high enough (up to sixth order for high-quality applications, i.e., 60–70-dB mirror frequency suppression); and in some cases, the center frequency must be tunable. A filter with these specifications cannot be integrated. These filters are realized with discrete components. They consist of capacitors and inductors which have to be tuned during production. The tunability of the center frequency is realized by means of a discrete varicap diode. These HF filters are expensive and vulnerable.

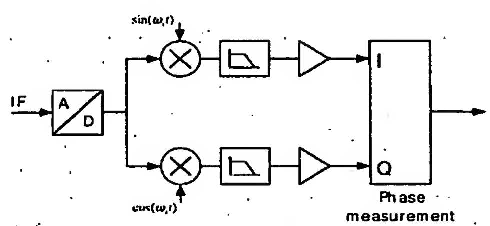

Fig. 3. Demodulation with a DSP of a signal situated on an IF.

Once the signal is downconverted to the IF, it has to be filtered further on in order to extract the wanted signal from its neighbors. This filter too must have a high Q (e.g., 50) and a high order (eighth or tenth order). Integrating these IF filters is also very hard. Although ever more analog integrated IF filters are published [4], [5], for most applications they are still not good enough, and ceramic resonators are still used instead. In [6]–[8], it is shown that the performance (i.e., achievable dynamic range over power consumption) of integrated active bandpass filters is intrinsically Q-times worse than the perfor-

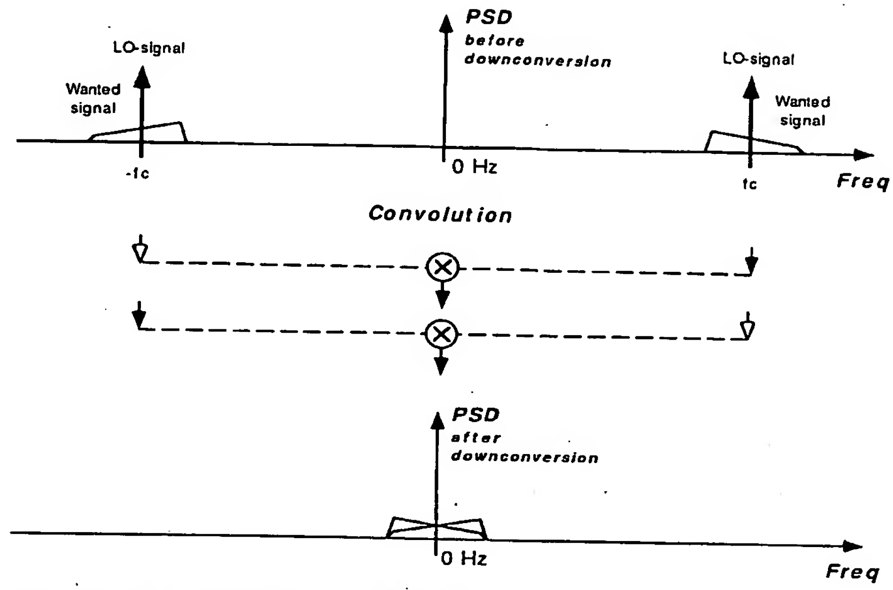

Fig. 4. Direct downconversion by mixing the RF signal with a single sine.

mance of a passive bandpass filter. This means that the discrete passive bandpass filters will always be a lot better than an active integrated version if one is concerned with power and area efficiency. The problem is that, compared to integrated filters, the use of these discrete components is very expensive.

Fig. 3 shows schematically how demodulation can be done with a DSP in an IF receiver. The IF signal is sampled and then downconverted in quadrature. Further demodulation is the same as in zero-IF receivers. The difference is that digitally the quadrature signal can be generated with a very high precision. Often the IF is too high for the A/D-converter, and then one or more extra IF stages are used. The final IF is often situated around 1 MHz [9].

B. Zero-IF Receivers

In zero-IF receivers, the wanted signal is directly downconverted to the baseband. The IF is chosen to be zero. In this case, the signal on the mirror frequency is the wanted signal itself. However, this does not eliminate the problem of the mirror frequency. Both signals—the wanted signal and the signal from the mirror frequency—although they are coming from the same carrier frequency, are not the same. In this case, they are each other's mirror image. Fig. 4 shows that this results in the lower and upper sideband being placed on top of each other in the baseband, which means that they become inseparable.

This problem is solved by doing the downconversion twice—once with a sine and once with a cosine. The topology of a zero-IF receiver is given in Fig. 5. In (2) and (3), the quadrature downconversion of the antenna signal $a(t)$ is calculated.

$$u_i(t) = a(t) \times \cos(2\pi f_c t + \varphi)$$
$$u_q(t) = a(t) \times \sin(2\pi f_c t + \varphi). \tag{2}$$

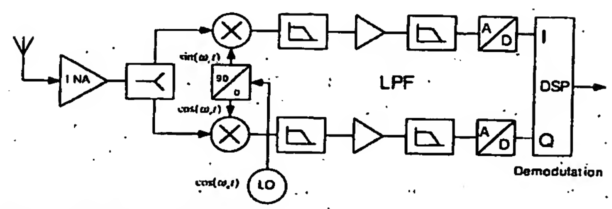

Fig. 5. The zero-IF receiver topology.

The original signal $m(t)$ can, in case of phase modulation, be found as the angle of the vector $[v_i(t), v_q(t)]$, a low-pass filtered version of $[u_i(t), u_q(t)]$.

$$m(t) = \arctan\left[\frac{v_q(t)}{v_i(t)}\right] + \varphi - \phi. \tag{3}$$

The demodulation in a DSP can be done by means of an angle measurement algorithm. The CORDIC algorithm is an example of such an algorithm [10].

The advantages of zero-IF receivers are obvious. There is no need for a high-Q tunable bandpass filter. In most designs, a broadband HF filter which requires no tuning or adjusting is used to reduce the dynamic range requirements for the downconversion part and prevent mixing of the RF signal with harmonic components of the LO signal. The low-pass filters can easily be realized as analog integrated filters.

The precision with which both demodulation paths (I and Q) can be matched determines how good the mirrored signal can be suppressed. The specifications on mirror suppression are not as severe for a zero-IF receiver as they are for an IF receiver. In an IF receiver, extra suppression is needed because the signal on the mirror frequency can be bigger than the wanted signal. However, a suppression of 40 dB is still needed in a good quality zero-IF receiver. The zero-IF receiver

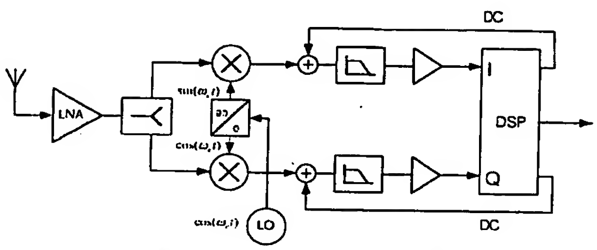

Fig. 6. A zero-IF receiver with a DC compensation feedback path.

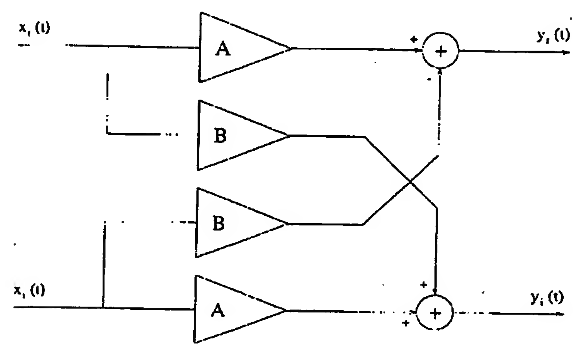

Fig. 7. Block scheme of the amplification of a complex signal with a complex constant.

is thus very sensitive to matching and to phase and amplitude errors in the quadrature oscillator.

Another problem in zero-IF receivers is the DC-offset which is created during the downconversion. It is mainly a result of the crosstalk between the RF and LO inputs of the mixer. The multiplication of the LO signal with itself gives a DC (or almost DC) signal. This DC-offset is superimposed on the wanted signal in the baseband. It can only be removed be means of very long time constants (at least one-tenth of a second), and it always results in the loss of a part of the wanted signal. This has an effect that is comparable to distortion [11]. The distortion will be lower and of an acceptable level if the time constant is longer (e.g., 1 s), but this long time constant makes the settling time of the complete receiver system too long, and these time constants cannot be analog integrated. For instance, at each change of the carrier frequency, the receiver would have to settle for at least 1 s when a highpass filter of 1 Hz is used. The principle of zero-IF receivers has been known for years. However, it is the DC-offset problem that has kept the zero-IF receiver from use in practical applications. It is only with the introduction of DSP's that this problem has become controllable for system with lower quality specifications (i.e., based on digital signals) [12]. In the DSP a complex nonlinear scheme can be used to determine the DC-level dynamically. This value can then be fed back into the analog part. This is shown in Fig. 6. In this way, saturation of the low-pass filters is prevented and the distortion is kept acceptable.

The crosstalk between LO and RF not only results in a DC voltage; the multiplication of the RF signal with itself results in a broadband baseband signal (twice the bandwidth of the RF signal). A considerable part of the power of this signal is situated in the lower baseband. This is an unwanted effect because this means that this signal is also superimposed on the wanted baseband signal; and once this is done, they cannot be separated anymore.

C. Low-IF Receivers

The development of the low-IF topologies starts from the perception that it must be possible to realize a receiver which combines the advantages of both known receiver types. If one would use two downconversion paths in an IF receiver, all required information for the separation of the wanted signal from the mirror signal would still be available in the two IF signals, as it is in a zero-IF receiver. So in this way it must be possible to postpone the mirror suppression from the HF part to the IF part. A high-quality HF filter would not be necessary anymore. A broadband HF filter, the same as for zero-IF receivers, would be satisfactory. The high IF, necessary to make the HF filter realizable, can now be replaced by a low IF (e.g., one or two times the bandwidth of the wanted signal) and the IF filter becomes now a set of two low Q ($Q = 1$ or 2) bandpass filters whose integration is as easy as the integration of the low-pass filters for a zero-IF receiver.

Although the concept of the low-IF receiver has been proposed in the previous paragraph, it is still not clear how such a receiver can be realized. First, it must be determined how the information available in both IF signals can be used to separate the wanted signal from the mirror signal. Therefore, in the next section of this paper, the complex signal technique is introduced. This is an analysis and synthesis technique which is used in different fields of digital signal processing [13], [14]. Its key concept is the use of multipath signal processing systems which are analyzed as single path systems. The quality of this multipath signal processing depends heavily on how well the operations in each path are matched with each other. In digital systems, this matching can be perfect; in analog systems, a perfect matching is impossible. In this paper, the complex signal technique will be used to analyze the analog zero-IF topology—an obvious multipath system. The limitations of analog integrated systems which use complex signals will be analyzed; and starting from this discussion, the low-IF receiver topologies will be synthesized. A low-IF receiver can be highly integrated and exhibit high performance at the same time.

III. Systems with Complex Signals

A. Polyphase and Complex Signals

A polyphase signal is, by definition, a vector of independent signals. In an n-phase system, these vectors are n-dimensional.

$$\begin{aligned} u(t) &= [u_1(t), u_2(t), \cdots, u_n(t)] \\ U(j\omega) &= [U_1(j\omega), U_2(j\omega), \cdots, U_n(j\omega)]. \end{aligned} \tag{4}$$

These polyphase signals can take any form, and they are therefore not very special. It becomes interesting when they are described as a sum of sequences on which operations

are defined which have sequence specific responses. In the following equations, this is worked out for the special case of two-phase signals. The notation is different from (4). Here the complex notation is used for the two-dimensional vectors and from now on two-phase signals are called complex signals. The j makes sure that the r and i component can be independent components of $u(t)$, and in this way the complex signal, which appears to be one signal, is two-dimensional.

$$\begin{aligned} u(t) &= u_r(t) + j u_i(t) \\ U(j\omega) &= U_r(j\omega) + j U_i(j\omega). \end{aligned} \tag{5}$$

Every frequency component of $u(t)$ can be written as a sum of two sequences. The two sequences of a real signal $[u(t) = u_r(t)]$ always have the same amplitude and the opposite phase.

$$\begin{aligned} &A(\omega) \cdot \cos[\omega t + \varphi(\omega)] \\ &\quad = \frac{A(\omega)}{2} \cdot \{\cos[\omega t + \varphi(\omega)] + j \cdot \sin[\omega t + \varphi(\omega)]\} \\ &\qquad + \frac{A(\omega)}{2} \cdot \{\cos[\omega t + \varphi(\omega)] - j \cdot \sin[\omega t + \varphi(\omega)]\}. \end{aligned} \tag{6}$$

The first sequence has only a positive frequency component, the second only a negative.

$$\begin{aligned} &A(\omega) \cdot \{\cos[\omega t + \varphi(\omega)] + j \cdot \sin[\omega t + \varphi(\omega)]\} \\ &\quad = A(\omega) \cdot e^{j\varphi(\omega)} \cdot e^{j\omega t} \\ &A(\omega) \cdot \{\cos[\omega t + \varphi(\omega)] - j \cdot \sin[\omega t + \varphi(\omega)]\} \\ &\quad = A(\omega) \cdot e^{-j\varphi(\omega)} \cdot e^{-j\omega t}. \end{aligned} \tag{7}$$

In this way, any complex signal, i.e., any ordered set of two independent signals, can be represented as a sum of positive and negative frequency components which in turn are also totally independent. Making a difference between positive and negative frequencies becomes an advantage if one can realize operators which make a difference between positive and negative frequency components.

B. Operations on Complex Signals

The use of complex signals becomes interesting once there are some specific operations defined on them which can be realized in an analog integrated way. The sum of two complex signals is a first and trivial operations. Each component of the sum signal is equal to the sum of the respective components of the signals that are summed. This does of course also mean that frequency components are not transformed to new positions by this operation.

$$\begin{aligned} z(t) &= x(t) + y(t) = [x_r(t) + j x_i(t)] + [y_r(t) + j y_i(t)] \\ &= [x_r(t) + y_r(t)] + j \cdot [x_i(t) + y_i(t)]. \end{aligned} \tag{8}$$

The complex multiplication, defined by $j^2 = -1$, introduces some more interesting operations. The multiplication of a complex signal with a complex constant is not interesting as such. It is, however, a useful building in the synthesis of complex filters.

$$\begin{aligned} y(t) &= Z \cdot x(t) = (Z_r + j Z_i) \cdot [x_r(t) + j x_i(t)] \\ &= [Z_r \cdot x_r(t) - Z_i \cdot x_i(t)] + j[Z_r \cdot x_i(t) + Z_i \cdot x_r(t)]. \end{aligned} \tag{9}$$

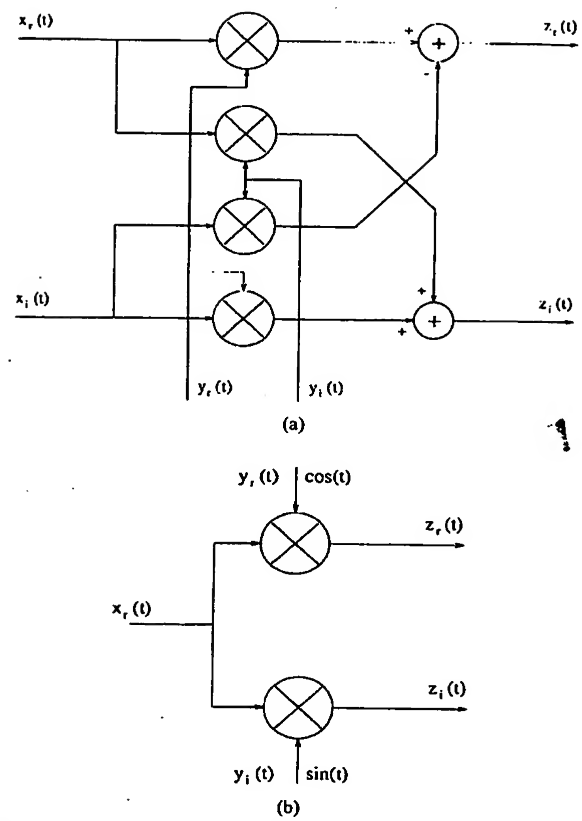

Fig. 8. (a) The multiplication of a complex signal with a complex signal. (b) The special case of the multiplication of a real signal with a complex signal.

Fig. 7 shows how this operation can be realized (with $Z = A + jB$). It is nothing more than the direct implementation of (9). The multiplication of two complex signals is a very important operation in receiver design.

$$\begin{aligned} z(t) &= y(t) \cdot x(t) = [y_r(t) + j y_i(t)] \cdot [x_r(t) + j x_i(t)] \\ &= [y_r(t) \cdot x_r(t) - y_i(t) \cdot x_i(t)] \\ &\quad + j[y_r(t) \cdot x_i(t) + y_i(t) \cdot x_r(t)] \end{aligned} \tag{10}$$

$$\begin{aligned} Z(j\omega) &= Y(j\omega) \otimes X(j\omega) \\ &= [Y_r(j\omega) + j Y_i(j\omega)] \otimes [X_r(j\omega) + j X_i(j\omega)] \\ &= [Y_r(j\omega) \otimes X_r(j\omega) - Y_i(j\omega) \otimes X_i(j\omega)] \\ &\quad + j[Y_r(j\omega) \otimes X_i(j\omega) + Y_i(j\omega) \otimes X_r(j\omega)]. \end{aligned} \tag{11}$$

The realization follows again from the worked out equations and is shown in Fig. 8(a). With this complex mixer it is possible to do a downconversion with only a positive or negative frequency. A sine has both, and this leads to the problem of the mirror frequency. The wanted signal is downconverted with the positive frequency component, and the mirror signal is downconverted with the negative frequency component. A special case is the downconversion of a real signal with a positive frequency. The scheme of Fig. 8(a) is reduced to

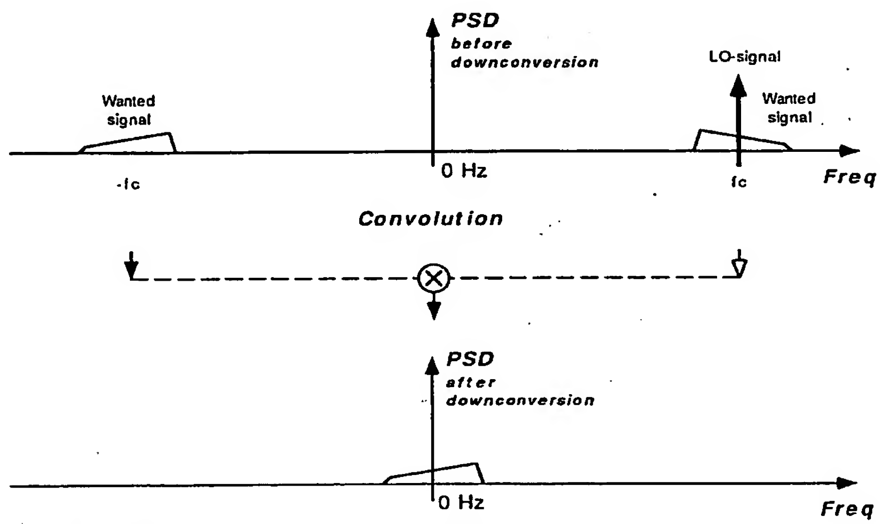

Fig. 9. Downconversion in a zero-IF receiver: mixing with a positive frequency.

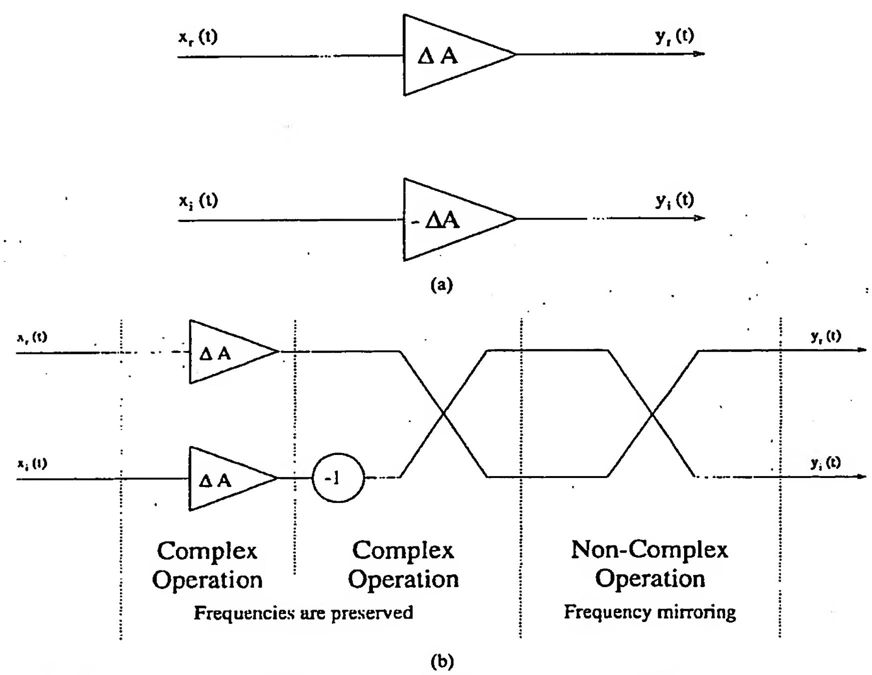

Fig. 10. (a) Mismatch between the amplification in the real and imaginary path. (b) Equivalent scheme.

Fig. 8(b), and it is obvious that this is the mixer which is used in zero-IF receivers. Fig. 9 shows the downconversion of a real signal with a single positive frequency—the downconversion which is used in a zero-IF receiver. Only the signal situated on the negative frequencies is downconverted, and in this way there is no superposition of the lower and upper sideband at baseband.

The most important operation on complex signals is filtering. A complex filter has a transfer characteristic in function of frequency which is sequence dependent. This means that the filtering of positive frequencies is different from the filtering of negative frequencies. Some positive frequencies may be passed, while the same negative frequencies are suppressed. This effect is very handy in receiver design, especially in those cases where the mirror signal is situated on the opposite frequencies of the wanted signal. Once the required transfer function is known, the filter can again be realized from the worked-out equation. The topology is the same as the one of

Fig. 7. In this case, the amplifiers are frequency dependent.

$$\begin{aligned} Y(j\omega) &= H(j\omega) \cdot X(j\omega) \\ &= [H_r(j\omega) + jH_i(j\omega)] \cdot [X_r(j\omega) + jK_i(j\omega)] \\ &= [H_r(j\omega) \cdot X_r(j\omega) - H_i(j\omega) \cdot X_i(j\omega)] \\ &\quad + j[H_r(j\omega) \cdot X_i(j\omega) + H_i(j\omega) \cdot X_r(j\omega)]. \end{aligned} \tag{12}$$

This synthesis technique—the realization of a complex filter with real filters—is, however, very costly. The realization of a first-order filter with a complex pole, for instance, requires the use of four real second-order filters. Direct synthesis is a more appropriate realization method. Direct synthesis is the realization of the complex transfer function with complex summators, amplifiers, and integrators. In the next section of this text, this will be worked out, and the implications of a practical realization will be discussed.

C. Imperfections in Complex Systems

Complex operators are made with pairs of real operators, amplifiers, mixers, and filters. The performance of the system in which these complex operators are used degrades when they are not perfectly matched. In digital systems, matching can be perfect; in analog integrated implementations, mismatch is unavoidable. An example of the effect of mismatch is given in Fig. 10. The amplification of a complex signal with a real constant A normally does not change the frequency distribution of the signal. The mismatched amplification of Fig. 10(a) can, for analysis purposes, be splitted in three different operations: a perfectly matched amplification, a complex multiplication with $-j$, and a switching of the real and imaginary part of the complex signal. This is depicted in Fig. 10(b). The first two operations are complex operations which preserve the frequency distribution. The third operation—the switching of the real and imaginary part—is not a linear operation which preserves the frequencies. Switching the vector components of a complex signal is equal to reversing the vector sequences. A positive sequence becomes a negative sequence and vice versa. In (13) this is written down for a positive and a negative frequency:

$$\begin{aligned} x_r(t) + j \cdot x_i(t) &\rightarrow x_i(t) + j \cdot x_r(t) \\ e^{j\omega_c t} = \cos(\omega_c t) + j \cdot \sin(\omega_c t) &\rightarrow \sin(\omega_c t) + j \cdot \cos(\omega_c t) \\ &= j \cdot e^{-j\omega_c t} \\ e^{-j\omega_c t} = \cos(\omega_c t) - j \cdot \sin(\omega_c t) &\rightarrow \sin(\omega_c t) - j \cdot \cos(\omega_c t) \\ &= -j \cdot e^{j\omega_c t}. \end{aligned} \tag{13}$$

Using this operator for the example of Fig. 10, the total transfer function of the amplifier with mismatch can be calculated. An applied frequency component will be magnified with a factor A, but it will also be mirrored to its opposite frequency and magnified with a factor ΔA. The ratio between the unwanted mirrored signal and the wanted original signal is thus $\Delta A/A$. An amplitude error of 1% gives a −40-dB crosstalk between the positive and negative frequencies.

The result of a phase error between the two signal paths is also an unwanted mirroring of signals to their opposite frequencies. Equation (14) gives the effect of a phase error

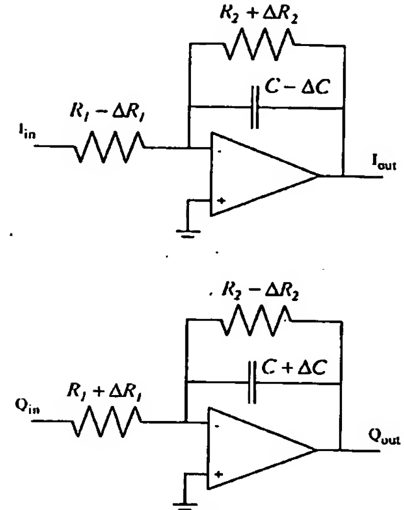

Fig. 11. Mismatch in a low-pass filter for complex signals.

on a positive frequency component.

$$\begin{aligned} e^{j\omega_c t} &= \cos(\omega_c t) + j \cdot \sin(\omega_c t) \\ &\rightarrow \cos(\omega_c t + \Delta\varphi) + j \cdot \sin(\omega_c t - \Delta\varphi) \\ &= \cos(\Delta\varphi) \cdot (e^{j\omega_c t} - j \cdot \tan(\Delta\varphi) \cdot e^{-j\omega_c t}). \end{aligned} \tag{14}$$

Apart from the wanted positive frequency, the total transfer function again generates also a negative and unwanted frequency component. The ratio between the unwanted and the wanted signal is $\tan(\Delta\varphi)$, which can be taken equal to $\Delta\varphi$ for small values of $\Delta\varphi$. This means that a phase error of 1° results in a −35-dB crosstalk between positive and negative frequencies.

The relationship between component matching and amplitude and phase errors is illustrated with an example. Fig. 11 gives the realization of an active-RC single pole low-pass filter for complex signals. All passive devices differ from their nominal values. There are three sources of unwanted frequency mirroring in this circuit. One is a constant amplitude error caused by the mismatch on R_1 and R_2. Equation (15) gives a calculation and approximation for this error.

$$\begin{aligned} \frac{\Delta A}{A} &= \left|\frac{A_r - A_i}{2A}\right| = \frac{\dfrac{R_2+\Delta R_2}{R_1-\Delta R_1} - \dfrac{R_2-\Delta R_2}{R_1+\Delta R_1}}{2\dfrac{R_2}{R_1}} \\ &= \frac{1}{2} \cdot \left(\frac{1+\Delta R_2/R_2}{1-\Delta R_1/R_1} - \frac{1-\Delta R_2/R_2}{1+\Delta R_1/R_1}\right) \\ &\approx \frac{1}{2} \cdot \left[\left(1+\frac{\Delta R_1}{R_1}\right)\cdot\left(1+\frac{\Delta R_2}{R_2}\right)\right. \\ &\qquad \left. - \left(1-\frac{\Delta R_1}{R_1}\right)\cdot\left(1-\frac{\Delta R_2}{R_2}\right)\right] \\ &\approx [\text{for all } \omega]\ \frac{\Delta R_1}{R_1} + \frac{\Delta R_2}{R_2}. \end{aligned} \tag{15}$$

The other two sources of unwanted frequency mirroring are the frequency dependent amplitude and phase error caused by the

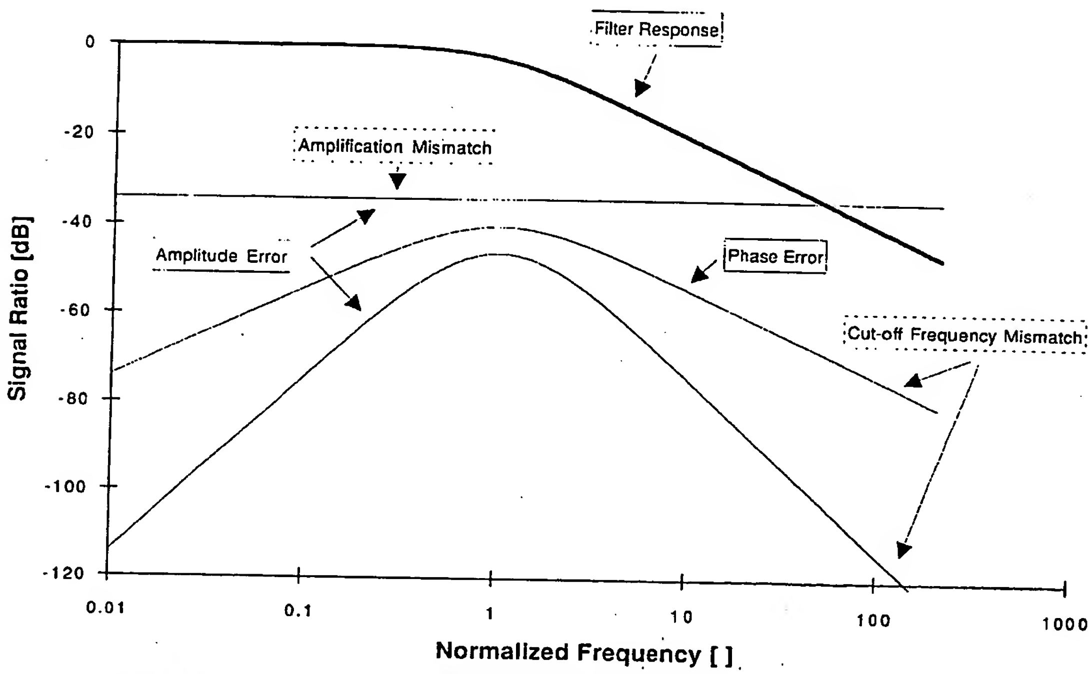

Fig. 12. Signal mirroring caused by mismatch in a low-pass filter for complex signals in function of frequency.

mismatch between the cutoff frequencies of the two low-pass filters. Equations (16) and (17) give these errors in function of frequency for a worst-case situation.

$$\begin{aligned}\tan(\Delta\varphi) &= \tan\left(\frac{\varphi_r - \varphi_i}{2}\right)\\ &= \tan\left\{\frac{1}{2}\cdot\arctan\left[\omega RC\cdot\left(1+\frac{\Delta R_2}{R_2}\right)\cdot\left(1+\frac{\Delta C}{C}\right)\right] - \frac{1}{2}\cdot\arctan\left[\omega RC\cdot\left(1-\frac{\Delta R_2}{R_2}\right)\cdot\left(1-\frac{\Delta C}{C}\right)\right]\right\}\\ &\approx [\text{for } \omega RC \text{ small}]\ \omega RC\cdot\left(\frac{\Delta R_2}{R_2}+\frac{\Delta C}{C}\right)\\ &\approx [\text{for} \omega RC = 1]\left(\frac{\Delta R_2}{2R_2}+\frac{\Delta C}{2C}\right).\end{aligned} \tag{17}$$

These three sources of mirroring are plotted versus frequency in Fig. 12. The curves of Fig. 12 are given for a 1% mismatch on each component. For (15)–(17) and Fig. 12, the worst-case analysis technique has been used for the simplicity of the equations. An insight into the total mirror error of a circuit, its mean value, and its variation can be obtained by doing a Monte Carlo simulation. The conclusions that can be drawn from Fig. 12 are the following: amplitude errors caused by pole position mismatch are small (so, capacitor mismatch does not result in amplitude errors); phase errors are caused by resistor and capacitor mismatch (they occur, however, only at the edge of the passband); the amplification mismatch is the main cause of signal mirroring errors in the passband (it completely depends on resistor matching).

The mirror error component that is caused by amplification mismatch can be corrected with the use of two matched AGC's

$$\begin{aligned}\frac{\Delta A}{A} &= \frac{\sqrt{\dfrac{1}{1+(\omega RC)^2\cdot(1+\Delta R_2/R_2)^2\cdot(1+\Delta C/C)^2}} - \sqrt{\dfrac{1}{1+(\omega RC)^2\cdot(1+\Delta R_2/R_2)^2\cdot(1-\Delta R_2/R_2)^2\cdot(1-\Delta C/C)^2}}}{\sqrt{\dfrac{1}{1+(\omega RC)^2}}}\\ &\approx \frac{\sqrt{\dfrac{1}{1+(\omega RC)^2\cdot(1+2\Delta R_2/R_2+2\Delta C/C)}} - \sqrt{\dfrac{1}{1+(\omega RC)^2\cdot(1-2\Delta R_2/R_2-2\Delta C/C)}}}{\sqrt{\dfrac{1}{1+(\omega RC)^2}}}\\ &\approx [\text{for } \omega RC = 1]\ \frac{\Delta R_1}{4R_1}+\frac{\Delta R_2}{4R_2}\end{aligned} \tag{16}$$

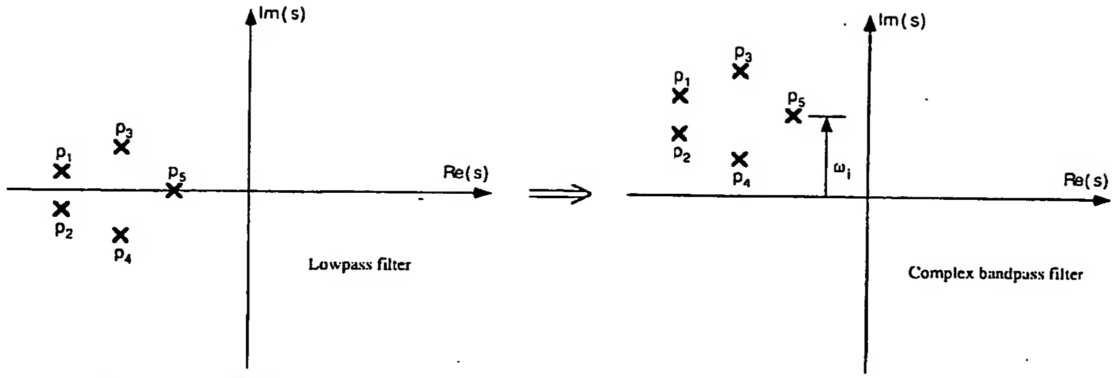

Fig. 13. S-plane representation of the frequency translation of a fifth-order low-pass filter.

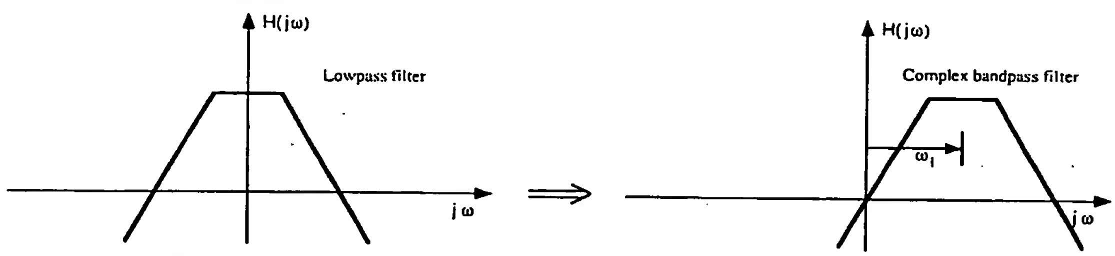

Fig. 14. The effect of a frequency translation on the transfer function.

in the real and imaginary signal path. However, this is only possible for circuits in which there is no physical coupling between the real and the imaginary path. This limits the usability of this technique to circuit topologies in which only real operations are performed on the complex signal. A zero-IF receiver is an example of such a circuit. The amplitude error is then of course determined by the matching between the AGC's; but in a DSP, AGC's can be implemented accurately. The other two types of mismatch error cannot be corrected. Although any phase and amplitude errors can be corrected theoretically, this is practically only possible when they are frequency independent or at least constant in the passband.

In the previous paragraph, the real first-order low-pass filter has been used as an example. The conclusions drawn from this example can be extended to higher order low-pass filters and amplifiers. Bandpass filters, oscillators, and mixers can also be analyzed in the same way. In bandpass filters, pole position matching is dominant. In the passband, the mirror error caused by amplification mismatch of bandpass filters differs not from low-pass filters, but the amplitude and phase errors caused by pole position mismatch are about Q-times worse. Quadrature oscillators often use an RC–CR circuit to obtain the 90° phase shift. The phase shift is a result of comparing the phase shift of a pole and a zero. Therefore, its mirror error highly depends on the matching of this pole and zero and will be dominantly a phase error.

IV. Analog Integrated Complex Filters

A. Complex Filters Synthesis

The transfer function of a real filter is a rational real polynomial function in $j\omega$. An active complex filter can have any rational complex polynomial function in $j\omega$ as transfer function. Most interesting are complex bandpass filters which have a passband only at either positive or negative frequencies [15], [16]. Their transfer function is found by frequency translating a low-pass filter.

$$H_{bp}(j\omega) = H_{lp}(j\omega - j\omega_c). \tag{18}$$

This translation remaps all poles and zero in the s-plane by moving them from their position centered around the zero axis to the center frequency f_c. The translation of poles and zeros in the s-plane is illustrated in Fig. 13. Fig. 14 shows the translation of the filter transfer function from low-pass to bandpass. The translation of a single pole is given in (19).

$$\begin{aligned} H_{lp}(j\omega) &= \frac{1}{1 + j\omega/\omega_o} \\ H_{bp}(j\omega) &= \frac{1}{1 - j\omega_c/\omega_o + j\omega/\omega_o} \\ &= \frac{1}{1 - 2jQ + j\omega/\omega_o}. \end{aligned} \tag{19}$$

A single complex pole cannot be realized with a real filter. Only complex pole pairs can be realized. The result of (19) is a single complex pole. The translated version of a single complex pole is also given with (19). The complex part must just be added to or subtracted from the complex term $2jQ$. Consequently, with this building block for a single complex pole, any translated low-pass filter transfer function can be realized. Fig. 13 depicts this frequency translation operation in the s-plane for a fifth-order low-pass Butterworth filter. After the translation, the poles are not compensated anymore. This is only possible in a complex filter.

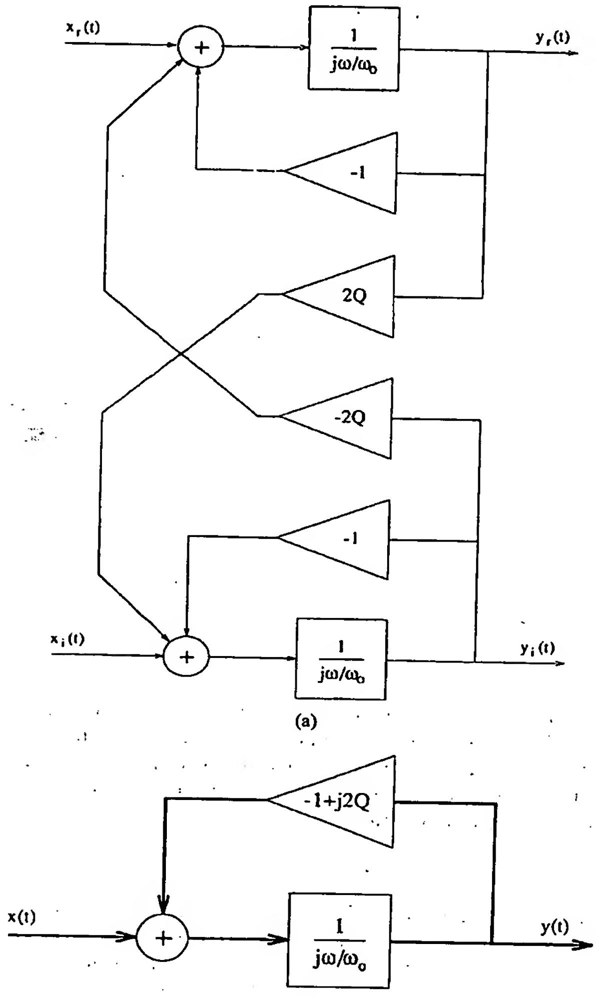

Fig. 15. Block scheme for the realization of a single complex pole. (a) The full notation and (b) the compact notation.

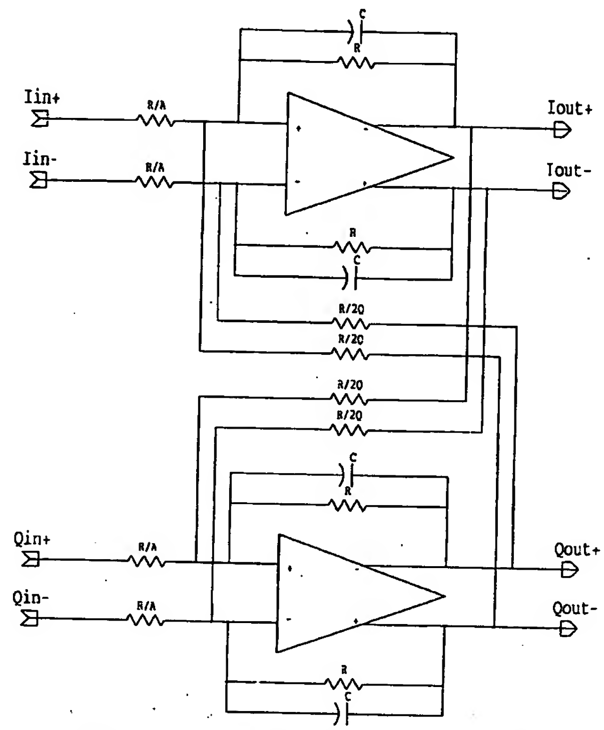

Fig. 16. Realization of a single complex pole with the active-RC filter technique.

The realization of $H_{bp}(j\omega)$ for a single pole is given in Fig. 15. It is nothing more than the direct synthesis of the transfer function. Fig. 15(a) is the full block schematic with building blocks for real signals. Fig. 15(b) is the same block schematic, but given in a more compact notation which will be used further on throughout this text. A thick arrow stands for a complex signal and it actually represents a bundle of signals (in all our examples, it represents either 2 or 4 signals). The building blocks can stand for complete complex operators. The realization of a translated pole via direct synthesis requires the use of two integrators. This is equal to what a real low-pass filter for complex signals requires, and it is only one-half of what a real bandpass filter for complex signals requires. This makes complex bandpass filter very cost efficient.

Analog integration of a complex filter can be done with all available implementation techniques for filters, as there are active-RC, OTA-C, OTA-RC, MOSFET-C, and Switched-Capacitor. Which technique is most appropriate depends on specifications imposed by the application. Fig. 16 shows an active-RC realization of a complex pole. In Fig. 17, the transfer function of a translated fifth-order low-pass Butterworth filter is given for positive and negative frequencies. The filter has a bandwidth of 220 kHz, and the center frequency is 250 kHz.

B. Sensitivity to Mismatch

In analog integrated implementations, which are different from digital implementations, perfect matching of components like resistors and capacitors is not possible. It is therefore important to choose an implementation technique that depends on components which have a good matching. For the active-RC realization, resistor and capacitor matching is important. Equations (20), (21), and (22) give approximate expressions for the errors at the edges of the passband of the translated low-pass filter of Fig. 15. This is only for frequency mirroring from negative to positive frequencies. The mirroring from wanted to unwanted frequencies is not an unwanted effect.

$$\frac{\Delta A}{A} = \frac{\Delta R}{R} + \frac{\Delta R}{2Q \cdot R} \quad \text{(amplification mismatch)} \quad (20)$$

$$\frac{\Delta A}{A} = \frac{\Delta R}{8R} + \frac{\Delta C}{8C} \quad \text{(pole position mismatch)} \quad (21)$$

$$\Delta\varphi = \frac{\Delta R}{4R} + \frac{\Delta C}{4C} \quad \text{(pole position mismatch).} \quad (22)$$

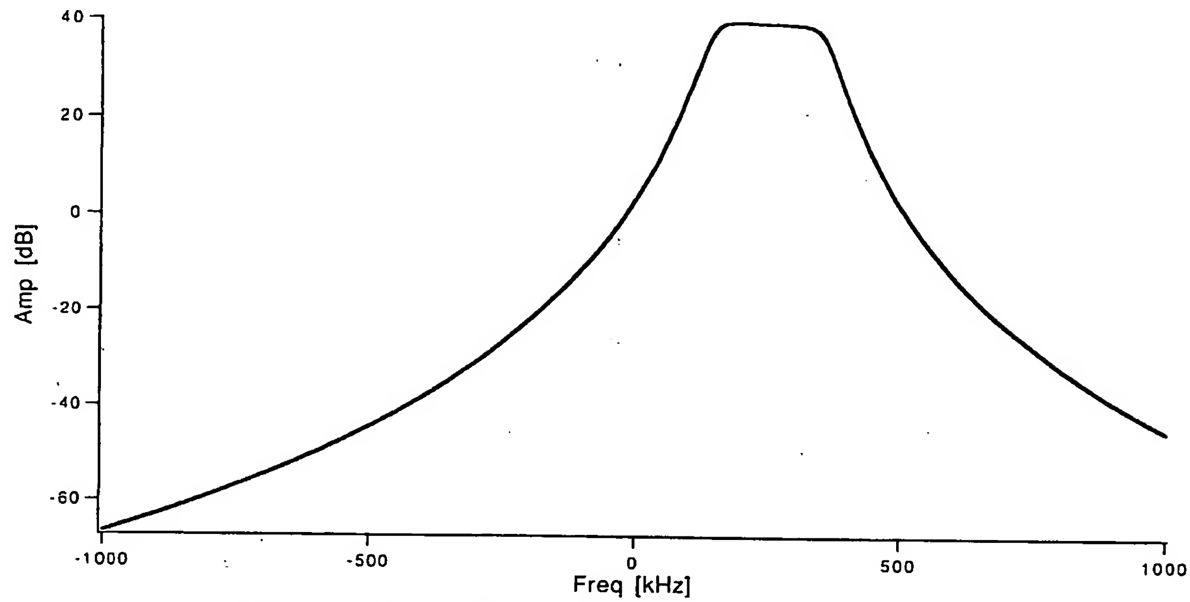

Fig. 17. Transfer function of a frequency translated fifth-order low-pass Butterworth filter.

Fig. 18 gives the crosstalk from negative to positive frequencies for the translated fifth-order Butterworth filter. This crosstalk has been simulated with the Monte Carlo simulation method assuming a random mismatch of 0.2% on every resistor and 0.1% on every capacitor. The crosstalk is equal to −60 dB and depends mainly on the mismatch between the input resistors of the first filter stage.

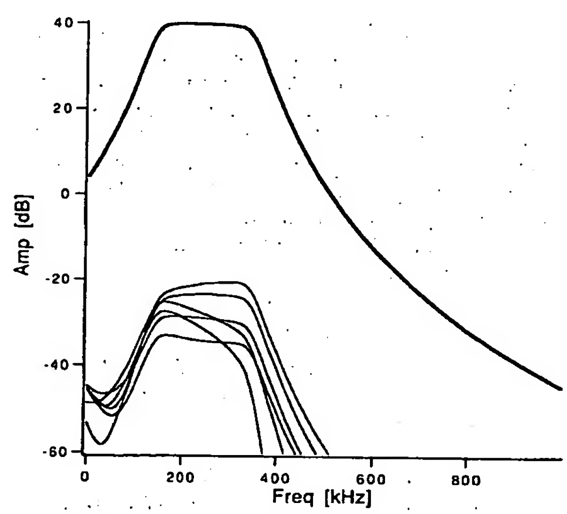

Fig. 18. Monte Carlo simulation of the signal crosstalk from negative to positive frequencies for the translated fifth-order Butterworth filter.

V. Receivers with Complex Signal Structures

A. Analysis of the Zero-IF Receiver

The obvious example of a receiver topology that uses complex signals is the zero-IF receiver. The topology of a zero-IF receiver is given, in compact notation, in Fig. 19. The RF signal is broadband filtered and then downconverted with a single positive frequency equal to the carrier frequency. The mixing with a positive frequency means that only the negative frequency components of the wanted signal are downconverted and that no mirrored signal is found in the baseband. This has already been illustrated with Fig. 7. After low-pass filtering, the signal is ready for sampling and demodulation.

B. The Low-IF Receiver using a Complex Filter

Fig. 20 shows how the quadrature downconversion mixer can also be used with a LO signal slightly lower then the carrier frequency. The wanted and the mirror signal are downconverted to the IF. However, both signals are not superimposed on each other. The wanted signal is situated at negative frequencies, while the mirror signal is situated at positive frequencies. This receiver topology can be called a low-IF receiver. The IF can be low (a few hundred kilohertz, i.e., one or two times the signal bandwidth), because there is no need for any HF mirror signal suppression.

Further downconversion of the IF signal can be done in two ways, both are shown in the full and compact notation in Figs. 21 and 22. In Fig. 21, the mirror signal is first suppressed with a complex bandpass, i.e., shifted low-pass filter, centered around the IF frequency. The final downconversion of the wanted signal to the baseband can then be done by multiplication with a sine. After the LF complex bandpass filter, the wanted signal is separated from its neighbors and is ready for sampling. The required dynamic range of the A/D-converter (number of bits) is the same as for a zero-IF receiver. The same signal is applied, only the center frequency is different. Digital applications require about 6–8 bits, high

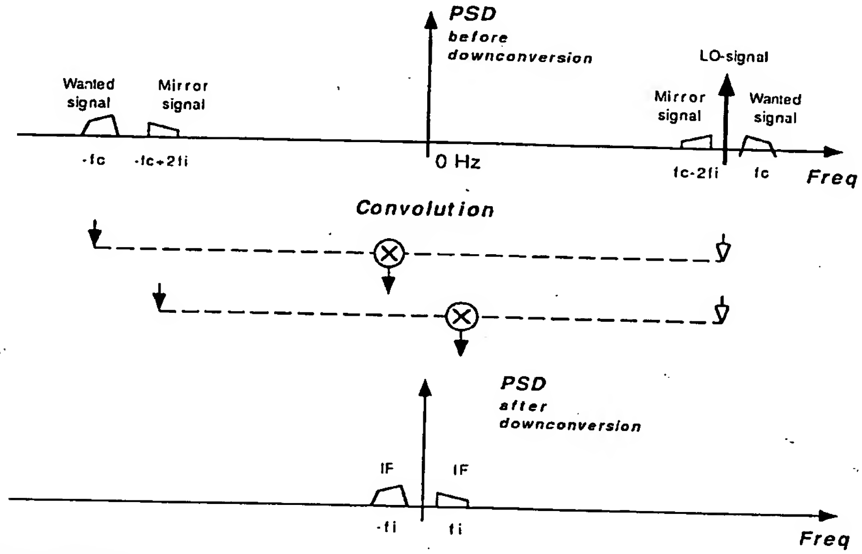

Fig. 19. Downconversion to a low IF.

quality continuous-time applications will require 10 bits. By making the sample frequency, f_s, equal to the IF, the sampling automatically performs the downconversion to the baseband. This means that the speed of the A/D-converter can also be the same as in zero-IF receivers (for the same suppression of aliasing signals with the same, but translated, LF filter). A 400-kHz IF and f_s are suitable values for a signal with a 200-kHz bandwidth (a 100 kHz baseband bandwidth). The same signal can be equally good downconverted with an IF of 200 kHz and an f_s of 400 kHz. In this configuration, every second sample has to be inverted.

C. The Low-IF Receiver Using Real Filters

Fig. 22 shows the alternative approach to the final downconversion. A real bandpass filter is used to separate both the wanted and the mirror signal from their neighbors. Downconversion must be done with a positive frequency equal to the IF. Downconversion with a single sine would bring the mirror signal also to the baseband. After downconversion, a low-pass filter makes sure that the mirror signal, situated at $2 \cdot f_i$ is suppressed sufficiently. Again, after the bandpass filter, the signal is ready for sampling. For easy downconversion, the sample rate has to be four times the IF (e.g., 800 kHz for a 200-kHz signal). This is twice the sample rate which is needed in the version with the complex bandpass filter, but this is no problem for today's A/D-converters. The big difference is found in the required number of bits. In this topology, not only the wanted signal, but the wanted signal together with the mirror signal, is sampled. The mirror signal can be 20–30 dB higher than the wanted signal, and in high-quality applications the wanted signal still has to be sampled with a 10 bit precision. Therefore, 4–5 extra bits are required, resulting in, for the high-quality applications, the need for a 16-bit A/D-converter. Low-cost digital applications will be satisfied with 12 bits, and this value can be further decreased by carefully choosing the position of the IF. The mirror frequency is best situated between two transmitter signals.

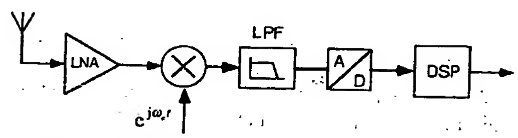

Fig. 20. The zero-IF receiver topology in compact notation.

Actually, the use of a real bandpass filter is not necessary. Using a 12–16-bit A/D-converter with a good antialiasing filter after the downconversion from RF is enough to make the topology of Fig. 22 work. The big advantage of the use of real filters is the fact that the frequency independent amplitude errors, which is the cause of a finite mirror rejection in the LF part, can be corrected by means of a simple AGC algorithm in the DSP.

D. Comparing Zero-IF and Low-IF Receivers

The advantages of a low-IF receiver compared to a zero-IF receiver are clear. There is absolutely no DC-problem simply because the wanted signal is not situated around DC. This also includes the problem of LO to RF crosstalk. The product of the LO multiplied with itself is situated around DC with a very small bandwidth (less than 1 kHz). The RF-to-LO crosstalk problem is reduced. The product of the RF signal multiplied with itself is broadband, but most of its power is still situated around DC in a band equal to the bandwidth of one transmitted signal (e.g., 200 kHz).

The disadvantage of low-IF receivers is that the suppression of the mirror signal must be higher. In zero-IF receivers, the mirror signal is the same as the wanted signal, and this means that a 40-dB suppression does result in an SNR of 40 dB for the wanted signal. In low-IF receivers, the mirror

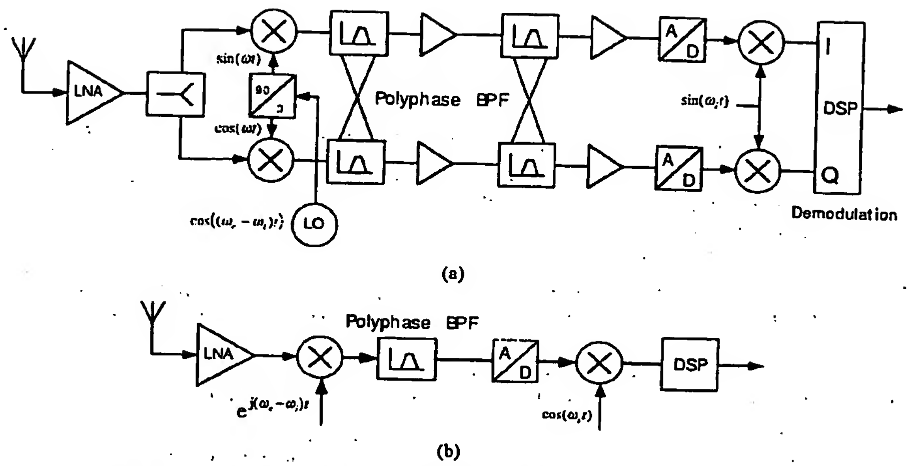

Fig. 21. The topology of a low-IF receiver which uses a complex bandpass filter in the LF part. (a) Full notation and (b) compact notation.

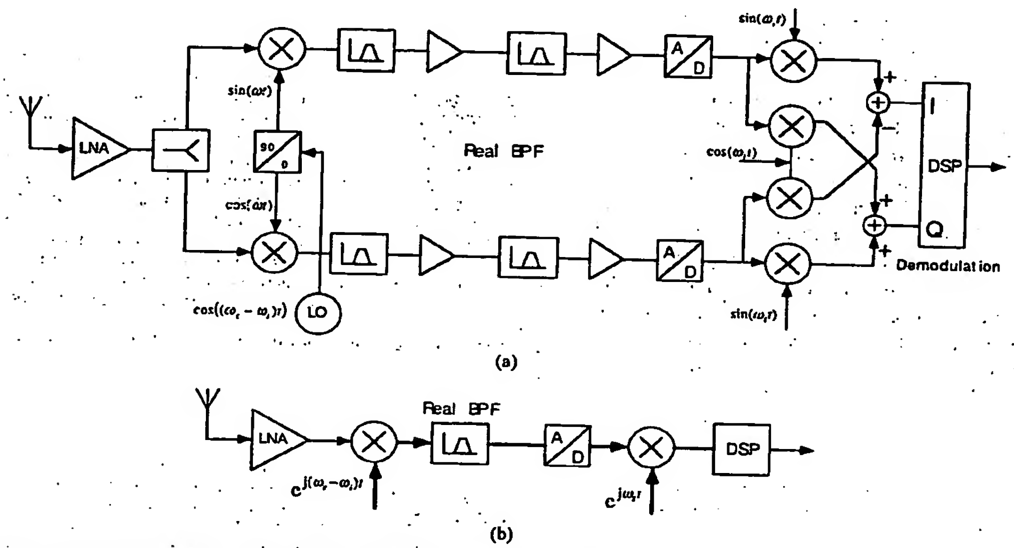

Fig. 22. The topology of a low-IF receiver which uses real filters in the LF part. (a) Full notation and (b) compact notation.

signal can be higher than the wanted signal. A mirror signal suppression of 70 dB is required for an SNR of 40 dB when the mirror signal can be 30 dB higher than the wanted signal. A careful choice of the IF, so that the mirror frequency is situated between two transmission channels, is important. In this way, the suppression specs can be lowered to 50–60 dB.

The achievable suppression is determined by the negative-to-positive frequency crosstalk of the complex building blocks, and this in its turn is determined by matching. A low-frequency complex bandpass filter with a negative-to-positive frequency crosstalk of less than −60 dB can be realized. Lower values, like −70 and −80 dB (required in high-quality applications), cannot be realized. The use of the low-IF topology with limited filtering (only antialiasing) and a 16-bit A/D-converter can offer this performance when a digital amplitude correction is applied. This strategy shifts the high specs from the analog part to the A/D-converter. As the performance of integrated A/D-converters improves rapidly, this is a good way to go. Even at this moment, a 16-bit 1-MHz A/D can already be realized at a reasonable cost of chip area and power consumption [17].

A matching problem that remains in all presented topologies is the phase error which is induced by the quadrature oscillator. Good matching is hard to realize at high frequencies and a phase error of 1° is equal to a mirror suppression as poor as −35 dB. It is this phase error which is the limit to the mirror suppression of todays zero-IF receivers. Further reduction of this phase error is vital, and a new technique has recently been proposed with which this phase error can be reduced to values which are significantly less than 1° [18].

VI. Conclusions

In this paper, the low-IF receiver has been examined as alternative for the IF and zero-IF receiver. Three new topologies have been presented. All three topologies use the same RF and

quadrature downconversion part as in a zero-IF receiver, but different from the zero-IF receiver are all three insensitive to DC offsets and LO to RF crosstalk. The three topologies have been synthesized by means of the complex signal method. With this technique, abstraction can be made of the multi-path approach which is used in zero-IF receivers and in the presented low-IF receivers. The complex signal technique has also been used to analyze the effects of mismatch between the paths in the different receiver topologies. Since low-IF receivers are insensitive to DC offsets, mismatch is the main remaining source of decreased performance because it results in a limited mirror signal suppression.

The first low-IF receiver topology which has been proposed is based on the replacement of the low-pass filters of a zero-IF receiver by a complex bandpass filter. Such a filter is a linear frequency translated version of the low-pass filters. However, it has only a passband for positive frequency components, and it is in this way that it can discriminate between the wanted signal and the mirror signal. The RF-part, the downconversion part, the A/D-converters, and the DSP for this low-IF receiver can be exactly the same as in the zero-IF receiver.

The second and third presented low-IF receiver use, respectivel, a bandpass filter and a slightly more broadband low-pass filter. In these receivers, the mirror signal suppression is not performed at the low IF, but in the DSP after a four-mixer further downconversion. The consequence is that these low-IF receivers require A/D-converters with a larger number of bits. Their advantage over the first low-IF receiver is that here the amplitude mismatches, which have been introduced in the analog part, can still be corrected in the DSP. The preferred version is the topology with the low-pass filters. It has the most simple structure of all three, and it is the least sensitive to phase errors.

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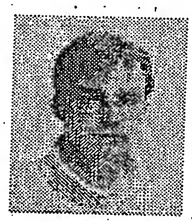

Jan Crols (S'90) was born in Turnhout, Belgium, in 1969. He received the M.Sc. degree in electrical and mechanical engineering in 1992 and the Ph.D. degree in electronics in 1997 from the Katholieke Universiteit Leuven, Heverlee, Belgium. His Ph.D. research was on the design of high performance fully integrated transceivers in CMOS technologies. For this work he obtained a fellowship of the Flemish Institute for Promotion of Scientific-Technological Research in Industry (IWT).

Currently, he is a Research Assistant at the ESAT-MICAS Laboratories of the Katholieke Universiteit Leuven. His research interests are in high-frequency analog integrated circuits for telecommunication.

Dr. Crols has been awarded the IEEE Solid-State Council Predoctoral Fellowship (1995–1996).

Michiel S. J. Steyaert (S'85–A'89–SM'92) was born in Aalst, Belgium, in 1959. He received the masters degree in electrical–mechanical engineering and the Ph.D. degree in electronics from the Katholieke Universiteit Leuven (K.U. Leuven), Heverlee, Belgium in 1983 and 1987, respectively.

From 1983 to 1986, he obtained an IWNOL fellowship (Belgian National Fundation for Industrial Research) which allowed him to work as a Research Assistant at the Laboratory ESAT at K.U. Leuven. In 1987, he was responsible for several industrial projects in the field of analog micropower circuits at the Laboratory ESAT as an IWONL Project Researcher. In 1988, he was a Visiting Assistant Professor at the University of California, Los Angeles. In 1989 he was appointed as a NFWO Research Associate and since 1992 a NFWO Senior Research Associate at the Laboratory/ESAT, K.U. Leuven, where he has been an Associate Professor since 1990. His current research interests are in high-performance and high-frequency analog integrated circuits for telecommunication systems and analog signal processing.

Prof. Steyaert received the 1990 European Solid-State Circuits Conference Best Paper Award, and the 1991 NFWO Alcatel-Bell-Telephone award for innovative work in integrated circuits for telecommunications.